PATENT
Attorney Docket No. 0064-0001RIS

IN THE UNITED STATES PATENT AND TRADEMARK OFFICE

In re Application of:	)
Ronald D. McCallister et al.	) Group Art Unit: 2611
Application No.: 10/718,507	) Examiner: Jean B. Corrielus
Filed: November 19, 2003	) Confirmation No.: 1244
For: Constrained-Envelope Digital-Communications Transmission System and Method Therefor	)

Commissioner for Patents
P.O. Box 1450
Alexandria, VA 22313-1450

Sir:

DECLARATION UNDER 37 C.F.R. § 1.132

I, J. Neil Birch, hereby make the following declaration:

1. I have been retained by Intersil Americas, Inc., assignee of the above-captioned reissue patent application.

2. I am an expert in communications systems, including OFDM radio systems.

3. This declaration supplements the declaration filed in the U.S. Patent and Trademark Office in the above-captioned patent application on May 18, 2006. My qualifications are set forth in that previous declaration and are not repeated herein.

4. I have read the "Inventor's Submission Under 37 C.F.R. 1.56" filed by Ronald D. McCallister (a named inventor of the above-captioned patent application) dated August 16, 2006 (the "August 16 Submission").

5. In the August 16 Submission, Mr. McCallister quotes on page 2 several sentences from the beginning of section II of the May et al. reference ("Reducing the Peak-to-Average Power Ratio in OFDM Radio Transmission Systems, "*Proceedings of the 1998 Vehicular Technology Conference*, May 18, 1998).

6. The portion quoted from May by Mr. McCallister includes the following:

> "This means that the signal is amplified linearly up to a maximal input amplitude A_o and larger amplitudes are limited to A_o, see Fig. [1]. Based on this assumption, we also model the amplifier as an ideal limiter with amplitude threshold A_o in this paper."

7. Mr. McCallister then states that it is his belief that the statement "the signal is amplified linearly" in the quoted passage "would be appreciated by a person of ordinary skill in the art as describing the claimed linearizer or linearizing limitations [of the claims of the above-captioned application]. In addition, Figure 1 of May, duplicated below, shows linearization up to a maximal input amplitude."

8. Mr. McCallister's representation of the teaching of May regarding use of a linear amplifier as set forth in the claims of the above-captioned application totally mischaracterizes the content of the May reference.

9. The amplifier described in May is modeled as an "ideal limiter" with normalized input and output amplitudes as shown in Fig. 1.

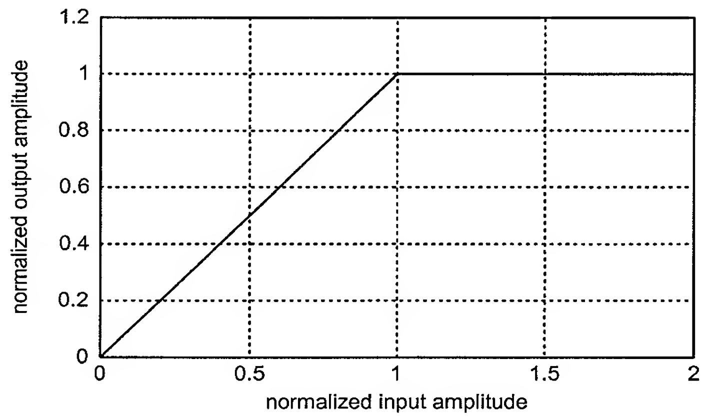

Fig. 1. Ideal limiter with normalized input and output amplitude, maximal input amplitude $A_0 = 1$

10. The amplifier described in May is a classic form of non-linear amplifier. Such amplifiers are also referred to as "clipping amplifiers" or "saturating amplifiers." See Spilker, *Digital Communications by Satellite*, 1977 (Exhibit A). Persons skilled in the art would understand that the amplifier modeled in May is a non-linear amplifier, contrary to Mr. McCallister's conclusion.

11. The May reference describes a theoretical system that manipulates an OFDM (Orthogonal Frequency Division Multiplexed) signal by introducing an additive correcting function that reduces amplitude peaks exceeding the clipping threshold A_o of the non-linear clipping amplifier. May at 2474, col. 2, lines 9-14.

12. May's additive correcting function k(t) is composed with the auxiliary function g(t) as set forth in the three equations at the bottom of page 2475, second column.

$$c(t) = s(t) + k(t) \quad where$$

$$k(t) = \sum_{n} A_n g(t - t_n)$$

$$A_n = -(|s(t_n)| - A_0)\frac{s(t_n)}{|s(t_n)|}$$

As noted (equation 3), A_n is expressed in terms of the clipping threshold A_o of the non-linear amplifier. Accordingly, May's correcting function is based on the characteristics, i.e., the clipping threshold, of the non-linear amplifier used in his system.

13. The constrained-envelope digital communications transmitter claimed in the above-captioned patent application is used with systems that receive quadrature phase-point signal streams and that operate with a substantially <u>linear</u> amplifier.

14. Persons skilled in the art would not find the subject matter of the claims of the above-captioned application requiring quadrature phase-point signal streams and a <u>linear</u> amplifier disclosed in, or obvious from, the teaching of May because the core teaching of May is to generate and apply a correcting signal that is tailored to the <u>non-linearity</u> of the amplifier used in the system.

15. I've reviewed the teaching of Dent patent 5,262,734. Dent describes a <u>linear</u> RF power amplifier 10 that produces intermodulation products at frequencies not present at the amplifier input. Dent uses a pre-distortion circuit and a distortion analyzer to introduce pre-distortions into the amplifier input to compensate for the intermodulation products. Dent col. 2, lines 7-15.

16. Dent's amplifier is not the "ideal limiter" required by the system modeled in May. Dent's amplifier is a <u>linear</u> amplifier. Dent col. 3, line 10. May's system would not

function as disclosed if the linear amplifier of Dent were substituted for May's non-linear amplifier.

17. May's correcting signal k(t) is generated only when the amplifier input exceeds the clipping threshold A_o. Dent's linear amplifier has no disclosed clipping threshold and receives only inputs which have been modified by the predistoring circuit so they are amplified in a linear manner. Dent col. 5, lines 36-48. Use of Dent's amplifier would thus render May's correcting signal generator non-operable, i.e., amplifier inputs would not exceed A_o.

18. The following supplements my declaration dated May 17, 2006, which was filed in the above-captioned application with the Amendment of May 30, 2006, on the matter of delaying the correcting signal k(t) in May.

19. May employs OFDM modulation. OFDM signals contain multiple subcarriers of different frequencies. For example, May's Fig. 3 shows a correcting function k(t) for an OFDM signal with 128 subcarriers. PSK or QAM modulation of the subcarriers is based on random distribution of data. The subcarriers are combined using inverse Fourier transformation. VanNee, *OFDM For Wireless Multimedia Communications*, 33-36 (Exhibit B).

20. May's continuous-time OFDM signal contains randomly occurring amplitude peaks attributable to the random data modulation patterns. See May Fig. 2. May's correcting signal k(t) is based on the Gaussian function g(t) of variance $\sigma_f^2 = 1/2\pi\sigma^2$. May at 2475, col. 2, lines 12-15. Hence, the peaks in May's correcting signal k(t) are not offset in time from the modulated signal s(t) by a constant interval.

21. Feeding May's modulated signal s(t) through a fixed delay element will not correctly align the amplitude peaks with the peaks in the correcting signal k(t). May provides no description of circuitry for correctly aligning the amplitude peaks of the modulated signal s(t) with the peaks in the correcting signal k(t). In my opinion a person of ordinary skill in the art could not implement May's system without undue experimentation.

I further declare that all statements made herein of my own knowledge are true and that all statements made on information and belief are believed to be true, and further, that these statements were made with the knowledge that willful false statements and the like so made are punishable by fine or imprisonment, or both, under Section 1001 of Title 18 of the United States Code, and that such willful false statements may jeopardize the validity of the application or any patent issuing thereon.

Dated: June 5, 2007

By: J Neil Birch
J. Neil Birch

EXHIBIT A

Digital Communications by Satellite

J. J. SPILKER, JR. Ph.D.
Chairman, Stanford Telecommunications, Inc.

PRENTICE-HALL, INC., *Englewood Cliffs, New Jersey*

Library of Congress Cataloging in Publication Data

Spilker, J J
Digital communications by satellite.

(Prentice-Hall information and system sciences series)
Bibliography: p. 625
Includes index.
1. Artificial satellites in telecommunication.
2. Data transmission systems. I. Title.
TK5104.S64 621.38'0423 75-43878
ISBN 0-13-214155-8

© 1977 by PRENTICE-HALL, INC.
Englewood Cliffs, New Jersey

All rights reserved. No part of this book
may be reproduced in any form or by any means
without permission in writing from the publisher.

10 9 8 7 6

Printed in the United States of America

PRENTICE-HALL INTERNATIONAL, INC., *London*
PRENTICE-HALL OF AUSTRALIA PTY. LIMITED, *Sydney*
PRENTICE-HALL OF CANADA, LTD., *Toronto*
PRENTICE-HALL OF INDIA PRIVATE LIMITED, *New Delhi*
PRENTICE-HALL OF JAPAN, INC., *Tokyo*
PRENTICE-HALL OF SOUTHEAST ASIA PTE. LTD., *Singapore*

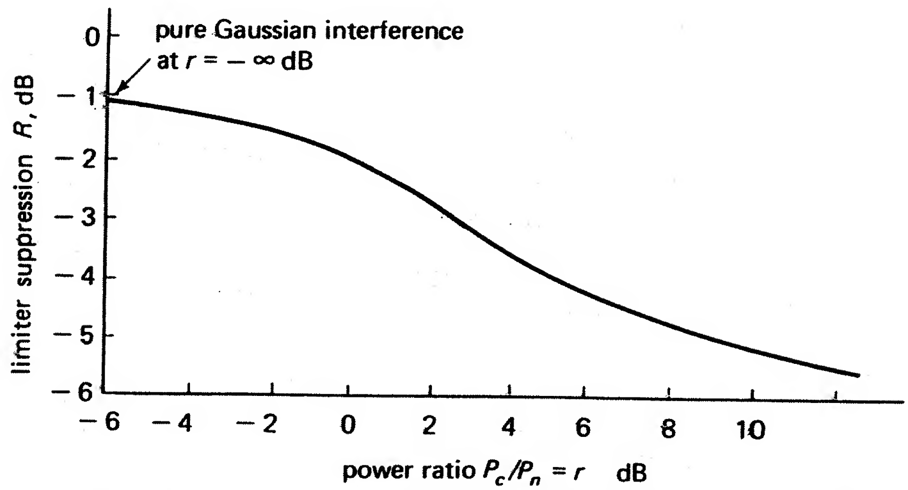

Fig. 9-13 Limiter suppression versus power ratio for a mixture of Gaussian and sinusoidal interference. The suppression approaches 6 dB as the interference becomes a pure sinusoid, $r = \infty$ dB.

9-6 DISTORTION CAUSED BY AMPLITUDE NONLINEARITIES

Signal suppression is only one of the effects caused by amplifier nonlinearities. One of the most important performance measures is the signal-to-distortion ratio—the ratio of individual signal power to the distortion power in the signal passband. We first determine the output signal-to-distortion spectral-density ratio versus the drive level. These results determine how much the transponder output power—power backoff—must be reduced to achieve a desired signal-to-distortion ratio for both sinusoidal and Gaussian inputs. As an intermediate step toward this calculation, the total signal power and total distortion power are calculated.

Piecewise Linear Bandpass Limiter

The transponder nonlinearity to be used here is the piecewise linear limiter, or clipping amplifier, shown in Fig. 9-14. The zonal filter passes only those frequency components in the fundamental frequency zone correspond-ing to the bandpass input. This model of a saturating amplifier is used in place of a hard limiter because it allows us to calculate the improvement resulting from power backoff. It is the simplest model of a saturating amplifier. Results for a hard limiter are obtained by letting the output be $y(t)/c$ and $c \to 0$.

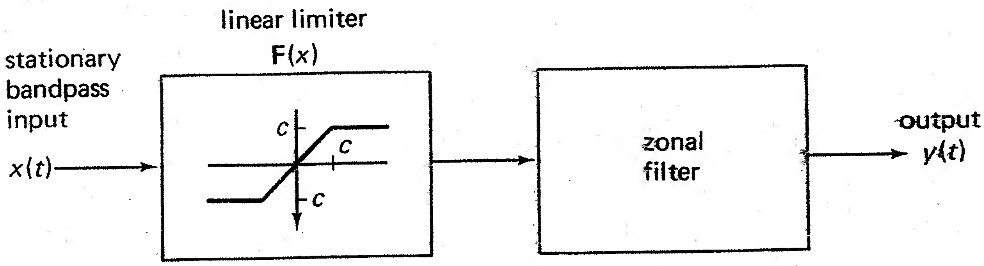

Fig. 9-14 Bandpass piecewise linear limiter model

Sinusoidal Input

As a first step, compute the variation in output power versus input power for the piecewise linear limiter. Let the input be a single sinusoid

$$x(t) = A\cos(\omega_0 t + \theta) = A\cos\phi \tag{9-29}$$

where $p(\phi) = \dfrac{1}{2\pi} \qquad |\phi| \leq \pi$

The output sinusoid amplitude in the fundamental zone is then

$$\begin{aligned} g(A) \triangleq B_1 &= \frac{1}{\pi}\int_0^{2\pi} f(A\cos\phi)\cos\phi\, d\phi \\ &= \frac{4}{\pi}\int_0^{\pi/2} f(A\cos\phi)\cos\phi\, d\phi \end{aligned} \tag{9-30}$$

For a linear limiter and a small amplitude input which produces no clipping, $A \leq c$, the system is linear and $B_1 = A$. For larger inputs $A \geq c$, where saturation occurs at least on the peaks of the sinusoid. The output is the sum of contributions from the limiting and linear regions:

$$\begin{aligned} B_1 &= \frac{4}{\pi}\left[\underbrace{c\int_0^{\cos^{-1} c/A} \cos\phi\, d\phi}_{\text{limiting region}} + \underbrace{\int_{\cos^{-1} c/A}^{\pi/2} A\cos^2\phi\, d\phi}_{\text{linear region}}\right] \\ &= \frac{4c}{\pi}\left\{\sin\phi\Big|_0^{\cos^{-1} c/A} + \frac{A}{2c}\left[1 + \frac{\sin 2\phi}{2}\right]_{\cos^{-1} c/A}^{\pi/2}\right\} \\ &= \frac{4c}{\pi}\left[\sqrt{1 - \left(\frac{c}{A}\right)^2} + \frac{A}{2c}\sin^{-1}\frac{c}{A} + \underbrace{\frac{A}{2c}\sin\phi\cos\phi\Big|_{\cos^{-1} c/A}^{\pi/2}}_{\frac{-A}{2c}\sqrt{1 - \left(\frac{c}{A}\right)^2}\frac{c}{A}}\right\} \end{aligned} \tag{9-31}$$

Thus, the output signal amplitude for this more general range of inputs is

$$B_1 = \frac{2c}{\pi}\left[\sqrt{1-\left(\frac{c}{A}\right)^2} + \frac{A}{c}\sin^{-1}\frac{c}{A}\right] \triangleq \gamma A \tag{9-32}$$

where γ is the equivalent gain. The power output at the fundamental frequency is $P_1 = B_1^2/2 \leq 8c^2/\pi^2$. The output power is plotted in Fig. 9-15 as a function of drive level $(A/c)^2$.

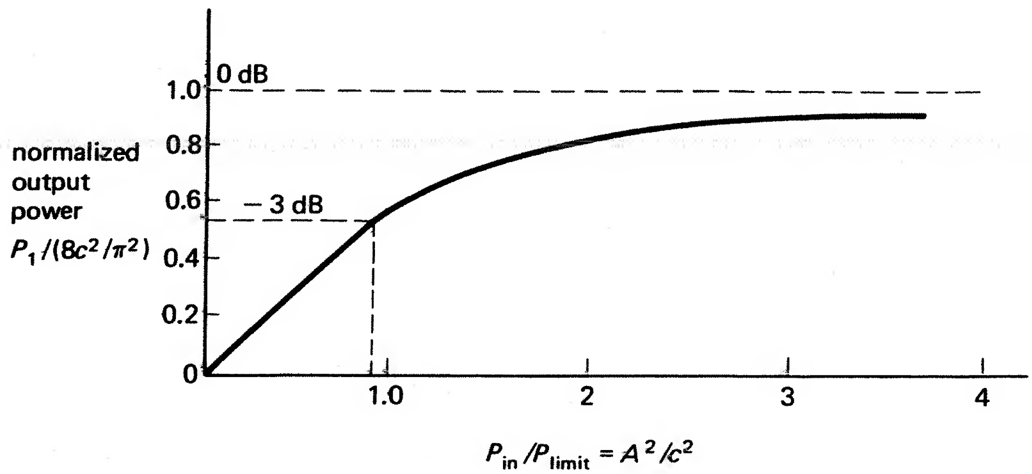

Fig. 9-15 Output versus input power for a piecewise linear limiter with a single sinusoidal input. The output power is normalized with respect to the maximum output power of the piecewise linear limiter.

An example of TWT transfer function (a "soft" limiter) is shown in Fig. 9-16. Both the output power and the intermodulation products are shown for two equal amplitude sine wave inputs.

Narrow-Band Gaussian Inputs

Consider a Gaussian input and compute the linear signal term and the intermodulation distortion components. Assume a bandpass input spectrum for the Gaussian input, as shown in Fig. 9-17. This spectrum can represent the sum of many frequency-multiplexed channels. If one channel is deleted, a notch would result in the input spectrum as shown. Intermodulation effects, caused by a nonlinear transponder amplifier or limiter, however, would partially fill in the notch in the amplifier output spectrum. The ratio of the intermodulation power in the notch to the signal power ordinarily in the notch gives the noise-power-ratio (NPR) or inverse SNR of that channel [Cahn, 1960, 53–59; Sunde, 1970, Chap. 8].

EXHIBIT B

OFDM for Wireless Multimedia Communications

Richard van Nee
Ramjee Prasad

Artech House
Boston • London

Library of Congress Cataloging-in-Publication Data
van Nee, Richard.
OFDM for wireless multimedia communications / Richard van Nee, Ramjee Prasad.
p. cm. — (Artech House universal personal communications library)
Includes bibliographical references and index.
ISBN 0-89006-530-6 (alk. paper)
1. Wireless communication systems. 2. Multimedia systems. 3. Multiplexing.
I. Prasad, Ramjee. II. Title. III. Series.

TK5103.2.N44 2000 99-052312
621.3845—dc21 CIP

British Library Cataloguing in Publication Data
van Nee, Richard
OFDM for wireless multimedia communications. —(Artech House
universal personal communications library)
1.Wireless communication systems 2. Multimedia systems
I. Title II. Prasad, Ramjee
621.3'82

ISBN 0-89006-530-6

Cover design by Igor Valdman

© 2000 Richard van Nee and Ramjee Prasad

All rights reserved. Printed and bound in the United States of America. No part of this book may be reproduced or utilized in any form or by any means, electronic or mechanical, including photocopying, recording, or by any information storage and retrieval system, without permission in writing from the authors.

All terms mentioned in this book that are known to be trademarks or service marks have been appropriately capitalized. Artech House cannot attest to the accuracy of this information. Use of a term in this book should not be regarded as affecting the validity of any trademark or service mark.

International Standard Book Number: 0-89006-530-6
Library of Congress Catalog Card Number: 99-052312

10 9 8 7 6 5 4

Chapter 2

OFDM Basics

2.1 INTRODUCTION

The basic principle of OFDM is to split a high-rate datastream into a number of lower rate streams that are transmitted simultaneously over a number of subcarriers. Because the symbol duration increases for the lower rate parallel subcarriers, the relative amount of dispersion in time caused by multipath delay spread is decreased. Intersymbol interference is eliminated almost completely by introducing a guard time in every OFDM symbol. In the guard time, the OFDM symbol is cyclically extended to avoid intercarrier interference. This whole process of generating an OFDM signal and the reasoning behind it are described in detail in sections 2.2 to 2.4.

In OFDM system design, a number of parameters are up for consideration, such as the number of subcarriers, guard time, symbol duration, subcarrier spacing, modulation type per subcarrier, and the type of forward error correction coding. The choice of parameters is influenced by system requirements such as available bandwidth, required bit rate, tolerable delay spread, and Doppler values. Some requirements are conflicting. For instance, to get a good delay spread tolerance, a large number of subcarriers with a small subcarrier spacing is desirable, but the opposite is true for a good tolerance against Doppler spread and phase noise. These design issues are discussed in Section 2.5. Section 2.6 gives an overview of OFDM signal processing functions, while Section 2. ends this chapter with a complexity comparison of OFDM versus single-carrier systems.

2.2 GENERATION OF SUBCARRIERS USING THE IFFT

An OFDM signal consists of a sum of subcarriers that are modulated by using *phase shift keying (PSK)* or *quadrature amplitude modulation (QAM)*. If d_i are the complex

QAM symbols, N_s is the number of subcarriers, T the symbol duration, and f_c the carrier frequency, then one OFDM symbol starting at $t = t_s$ can be written as

$$s(t) = \mathrm{Re}\left\{ \sum_{i=-\frac{N_s}{2}}^{\frac{N_s}{2}-1} d_{i+N_s/2} \exp\left(j2\pi\left(f_c - \frac{i+0.5}{T}\right)(t-t_s)\right) \right\} \quad , \; t_s \le t \le t_s + T \tag{2.1}$$

$$s(t) = 0\,, \quad t < t_s \;\wedge\; t > t_s + T$$

In the literature, often the equivalent complex baseband notation is used, which is given by (2.2). In this representation, the real and imaginary parts correspond to the in-phase and quadrature parts of the OFDM signal, which have to be multiplied by a cosine and sine of the desired carrier frequency to produce the final OFDM signal. Figure 2.1 shows the operation of the OFDM modulator in a block diagram.

$$s(t) = \sum_{i=-\frac{N_s}{2}}^{\frac{N_s}{2}-1} d_{i+N_s/2} \exp\left(j2\pi \frac{i}{T}(t-t_s)\right) \quad , \; t_s \le t \le t_s + T \tag{2.2}$$

$$s(t) = 0\,, \quad t < t_s \;\wedge\; t > t_s + T$$

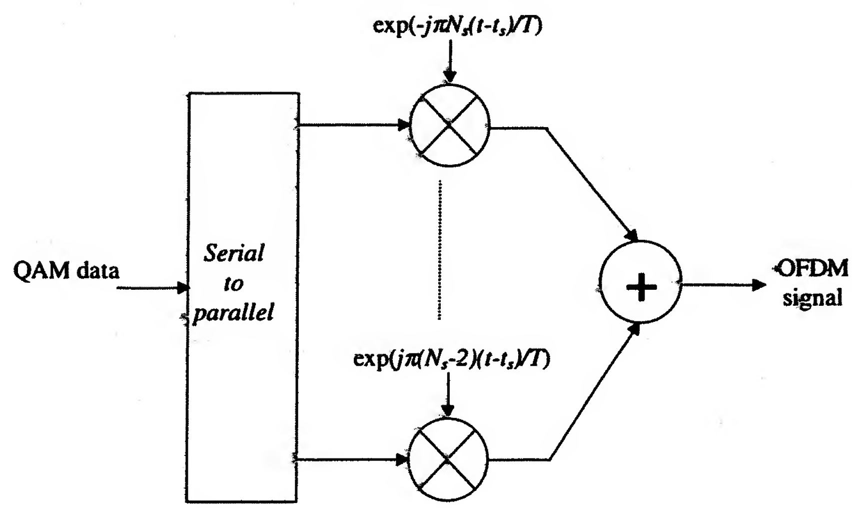

Figure 2.1 OFDM modulator.

As an example, Figure 2.2 shows four subcarriers from one OFDM signal. In this example, all subcarriers have the same phase and amplitude, but in practice the amplitudes and phases may be modulated differently for each subcarrier. Note that each subcarrier has exactly an integer number of cycles in the interval T, and the number of cycles between adjacent subcarriers differs by exactly one. This property accounts for

the orthogonality between the subcarriers. For instance, if the jth subcarrier from (2.2) is demodulated by downconverting the signal with a frequency of j/T and then integrating the signal over T seconds, the result is as written in (2.3). By looking at the intermediate result, it can be seen that a complex carrier is integrated over T seconds. For the demodulated subcarrier j, this integration gives the desired output $d_{j+N/2}$ (multiplied by a constant factor T), which is the QAM value for that particular subcarrier. For all other subcarriers, the integration is zero, because the frequency difference $(i-j)/T$ produces an integer number of cycles within the integration interval T, such that the integration result is always zero.

$$\begin{aligned}&\int_{t_s}^{t_s+T} \exp(-j2\pi \frac{j}{T}(t-t_s)) \sum_{i=-\frac{N_s}{2}}^{\frac{N_s}{2}-1} d_{i+N_s/2} \exp(j2\pi \frac{i}{T}(t-t_s))\, dt \\ &= \sum_{i=-\frac{N_s}{2}}^{\frac{N_s}{2}-1} d_{i+N_s/2} \int_{t_s}^{t_s+T} \exp(j2\pi \frac{i-j}{T}(t-t_s))dt = d_{j+N_s/2}T\end{aligned} \tag{2.3}$$

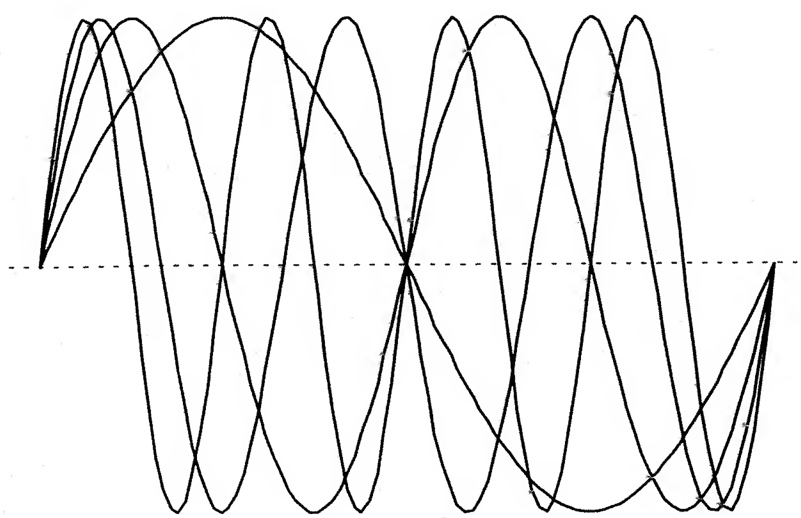

Figure 2.2 Example of four subcarriers within one OFDM symbol.

The orthogonality of the different OFDM subcarriers can also be demonstrated in another way. According to (2.1), each OFDM symbol contains subcarriers that are nonzero over a T-second interval. Hence, the spectrum of a single symbol is a convolution of a group of Dirac pulses located at the subcarrier frequencies with the spectrum of a square pulse that is one for a T-second period and zero otherwise. The amplitude spectrum of the square pulse is equal to $\mathrm{sinc}(\pi fT)$, which has zeros for all frequencies f that are an integer multiple of $1/T$. This effect is shown in Figure 2.2, which shows the overlapping sinc spectra of individual subcarriers. At the maximum of each subcarrier spectrum, all other subcarrier spectra are zero. Because an OFDM

receiver essentially calculates the spectrum values at those points that correspond to the maxima of individual subcarriers, it can demodulate each subcarrier free from any interference from the other subcarriers. Basically, Figure 2.3 shows that the OFDM spectrum fulfills Nyquist's criterium for an intersymbol interference free pulse shape. Notice that the pulse shape is present in the frequency domain and not in the time domain, for which the Nyquist criterium usually is applied. Therefore, instead of intersymbol interference (ISI), it is intercarrier interference (ICI) that is avoided by having the maximum of one subcarrier spectrum correspond to zero crossings of all the others.

The complex baseband OFDM signal as defined by (2.2) is in fact nothing more than the inverse Fourier transform of N_s QAM input symbols. The time discrete equivalent is the inverse discrete Fourier transform (IDFT), which is given by (2.4), where the time t is replaced by a sample number n. In practice, this transform can be implemented very efficiently by the inverse fast Fourier transform (IFFT). An N point IDFT requires a total of N^2 complex multiplications—which are actually only phase rotations. Of course, there are also additions necessary to do an IDFT, but since the hardware complexity of an adder is significantly lower than that of a multiplier or phase rotator, only the multiplications are used here for comparison. The IFFT drastically reduces the amount of calculations by exploiting the regularity of the operations in the IDFT. Using the radix-2 algorithm, an N-point IFFT requires only $(N/2)\cdot\log_2(N)$ complex multiplications [1]. For a 16-point transform, for instance, the difference is 256 multiplications for the IDFT versus 32 for the IFFT—a reduction by a factor of 8! This difference grows for larger numbers of subcarriers, as the IDFT complexity grows quadratically with N, while the IFFT complexity only grows slightly faster than linear.

$$s(n) = \sum_{i=0}^{N_s-1} d_i \exp(j2\pi \frac{in}{N}) \tag{2.4}$$

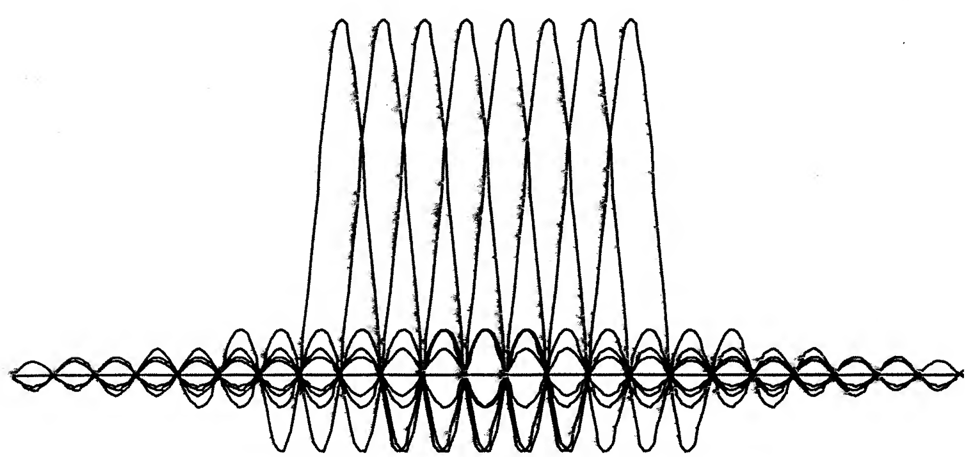

Figure 2.3 Spectra of individual subcarriers.